

Iterative Interference Cancellation for Near-Field XL-MIMO Systems with Line-of-Sight Channels

Nayyab Haider and Rodrigo C. de Lamare

Abstract—In extra-large multiple-input multiple-output (XL-MIMO) systems, near-field users experience spherical wavefront propagation, offering higher spatial resolution than far-field plane-wave models, but complicating signal detection and interference cancellation. Under line-of-sight conditions, we propose a weighted-rate successive interference cancellation (WR-SIC) scheme whose ordering metric exploits the enhanced spatial degrees of freedom of near-field propagation to prioritize users with stronger channels. An iterative detection and decoding (IDD) scheme with LDPC coding is then developed. Simulations in hybrid near- and far-field deployments show that the proposed WR-SIC and IDD schemes improve bit-error and sum rates over existing schemes.

Keywords—Near-field communications, XL-MIMO systems, line-of-sight channels, SIC, IDD schemes.

I. INTRODUCTION

Recent work on extra-large multiple-input multiple-output (XL-MIMO) systems has challenged traditional far-field assumptions and attracted a great deal of attention. In particular, when users operate within the Rayleigh distance, electromagnetic waves propagate as spherical rather than plane wavefronts [1], [2]. This near-field behavior creates spatial non-stationarity and visibility regions while offering enhanced spatial resolution and interference management capabilities. The coexistence of near-field and far-field users in practical deployments creates hybrid scenarios that present challenges for signal detection and interference cancellation [3].

In this context, successive interference cancellation (SIC) is fundamental for multiple-antenna systems, with existing ordering strategies including channel norm-based [4], [5] and SINR-based approaches [6], [7]. The V-BLAST architecture with zero-forcing SIC has been extensively analyzed [8], while recent work extends SIC to multiuser detection with MMSE processing [9], [10]. Furthermore, iterative detection and decoding (IDD) has shown significant gains when combined with SIC [10]–[12], while near-field channel modeling has gained attention due to unique Fresnel region characteristics [13], [14]. Recent advances address spatial non-stationarity and visibility regions [15]–[17], parametric channel estimation [18], and holographic MIMO communications [19]. Despite the progress in interference cancellation, IDD techniques, and near-field modeling, existing literature lacks integrated frameworks that exploit near-field spatial multiplexing characteristics. Comprehensive surveys on near-field XL-MIMO communications [20] highlight the need for novel signal

processing approaches that fully exploit the enhanced spatial degrees of freedom available in these systems.

In this work, we investigate iterative interference cancellation for extra-large multiple-input multiple-output (XL-MIMO) systems in near-field scenarios under line-of-sight (LoS) conditions. We develop an MMSE-based successive interference cancellation scheme with weighted-rate ordering, and integrate it with an iterative detection and decoding (IDD) scheme employing low-density parity-check (LDPC) codes. Our contributions are: (1) a weighted-rate ordering metric exploiting near-field rank enhancement; (2) integration with IDD for XL-MIMO; (3) performance analysis in hybrid near/far-field deployments with sum-rate and complexity analysis.

This work is organized as follows. Section II presents the system model. Section III introduces the SIC MMSE receiver with weighted-rate ordering and a sum-rate analysis, whereas Section IV details the proposed IDD scheme. Section V presents the results, while Section VI draws the conclusions.

II. SYSTEM MODEL

We consider the uplink of a near-field XL-MIMO system in which a base station (BS) equipped with M receive antennas serves K users, each equipped with N_U transmit antennas. The system operates at a carrier frequency of $f_c = 7.5$ GHz with a corresponding wavelength $\lambda = c/f_c$, where c is the speed of light. The BS employs a uniform linear array (ULA) with antenna spacing of $d = \lambda/2$. Each user transmits independent data streams from their antennas with equal power P_k/N_U per antenna (total power P_k), resulting in a total of $K \times N_U$ spatial streams that must be detected at the BS. No channel state information feedback to users is required: only the BS needs channel knowledge for detection and ordering.

This work focuses on LoS-dominant scenarios typical in near-field XL-MIMO deployments. Extension to multipath environments is left for future work. The electromagnetic radiation field surrounding antenna arrays can be divided into two distinct regions based on the propagation characteristics: the near-field region and the far-field region. The boundary between these regions is given by the Rayleigh distance, which depends on the array geometry and the operating frequency:

$$R = \frac{2D^2}{\lambda} \quad (1)$$

where D represents the maximum array aperture dimension. Users located at distances $d < R$ from the BS operate in the near-field region and experience spherical wavefront propagation, while users at distances $d \geq R$ are in the far-field region with plane wave characteristics.

A. Near-Field Channel Modeling

For users operating in the near-field LoS region, electromagnetic waves propagate as spherical wavefronts rather than the commonly assumed plane waves. This introduces significant complexity in channel modeling but also provides enhanced spatial discrimination capabilities. The near-field channel coefficient between the m -th transmit antenna of user k and the n -th receive antenna at the BS is modeled by

$$h_{n,m,k} = \sqrt{\beta_{n,m,k}} e^{-j\frac{2\pi}{\lambda}d_{n,m,k}} \quad (2)$$

Here $\beta_{n,m,k}$ denotes the Friis free-space power gain [21]. To state the power scaling explicitly, the channel coefficient is written with $\beta_{n,m,k}$ under a square root, so that the received power per antenna is $|h_{n,m,k}|^2 = \beta_{n,m,k}$, following the non-uniform spherical-wave model of [20].

$$\beta_{n,m,k} = \frac{\lambda^2}{(4\pi)^2 d_{n,m,k}^2} \quad (3)$$

with $d_{n,m,k}$ being the Euclidean distance between user antenna and each antenna of the BS that is calculated to capture the spherical wavefront characteristics as

$$d_{n,m,k} = \sqrt{\sum_{i \in \{x,y,z\}} (i_{tx,k,m} - i_{rx,n})^2} \quad (4)$$

where $i_{tx,k,m}$ and $i_{rx,n}$ represent the coordinates (x, y, or z) of the m -th transmit antenna of user k and the n -th receive antenna at the BS, respectively.

This distance calculation produces spatial non-stationarity, nonlinear phase relationships across antenna elements, and distance-dependent path loss variations, all of which provide enhanced spatial discrimination for interference cancellation.

B. Far-Field Channel Modeling

In contrast to the near-field scenario, users located in the far-field LoS region can be modeled using the plane wave approximation. This simplifies the channel modeling while maintaining reasonable accuracy for users at sufficient distances. The far-field channel coefficient is expressed by

$$h_{n,m,k} = \sqrt{\beta_k} e^{-j\frac{2\pi}{\lambda} \mathbf{k}_k \cdot \mathbf{r}_n} \quad (5)$$

where β_k is a distance-independent path loss coefficient for user k given by

$$\beta_k = \frac{\lambda^2}{(4\pi)^2 r_k^2} \quad (6)$$

Here, r_k denotes the average distance from user k to the BS, and the spatial variations across the array aperture are considered negligible. The wave vector $\mathbf{k}_k \in \mathbb{R}^{3 \times 1}$ encapsulates the directional information of the incident plane wave for user k and is defined in terms of the spherical coordinate system:

$$\mathbf{k}_k = \begin{bmatrix} \cos(\theta_{az,k}) \sin(\phi_{el,k}) \\ \sin(\theta_{az,k}) \sin(\phi_{el,k}) \\ \cos(\phi_{el,k}) \end{bmatrix} \quad (7)$$

where $\theta_{az,k}$ and $\phi_{el,k}$ are the azimuth and elevation angles of user k . The vector $\mathbf{r}_n \in \mathbb{R}^{3 \times 1}$ specifies the location

of the n -th receive antenna relative to a reference point, typically the array center. In the simulations, $\theta_{az,k}$ and $\phi_{el,k}$ are drawn independently and uniformly from $[0^\circ, 360^\circ)$ and $[0^\circ, 90^\circ]$, respectively, ensuring all users have distinct angular orientations with probability one [3]. All channel parameters are assumed perfectly known at the BS, consistent with the perfect CSI assumption [13].

C. Hybrid Network Configuration and Signal Model

We consider a hybrid network with K_{NF} near-field and K_{FF} far-field users, where $K_{NF} + K_{FF} = K$, applying spherical and plane wave models, respectively.

The channel matrix is constructed by concatenating channel vectors from all users:

$$\mathbf{H} = [\mathbf{h}_{1,1} \quad \mathbf{h}_{1,2} \quad \cdots \quad \mathbf{h}_{1,N_U} \quad \mathbf{h}_{2,1} \quad \cdots \quad \mathbf{h}_{K,N_U}] \quad (8)$$

where $\mathbf{h}_{k,n_u} \in \mathbb{C}^{M \times 1}$ represents the channel response from the n_u -th antenna of user k to all M BS antennas:

$$\mathbf{h}_{k,n_u} = [h_{1,n_u,k}, h_{2,n_u,k}, h_{3,n_u,k}, \dots, h_{M,n_u,k}]^T \quad (9)$$

The $M \times KN_U$ channel matrix $\mathbf{H}$ for M receive antennas and KN_U spatial streams can be partitioned into user-specific sub-matrices as given by

$$\mathbf{H} = [\mathbf{H}_1, \mathbf{H}_2, \mathbf{H}_3, \dots, \mathbf{H}_K] \quad (10)$$

where each $\mathbf{H}_k \in \mathbb{C}^{M \times N_U}$ contains responses from all antennas of user k as described by $\mathbf{H}_k = [\mathbf{h}_{k,1}, \mathbf{h}_{k,2}, \dots, \mathbf{h}_{k,N_U}]$.

The received signal represents the superposition of all transmitted streams with noise and is given by

$$\begin{aligned} \mathbf{y} &= \sum_{k=1}^{K_{NF}} \sum_{n_u=1}^{N_U} \mathbf{h}_{k,n_u} x_{k,n_u} + \sum_{k=K_{NF}+1}^K \sum_{n_u=1}^{N_U} \mathbf{h}_{k,n_u} x_{k,n_u} + \mathbf{n} \\ &= \sum_{k=1}^{K_{NF}} \mathbf{H}_k \mathbf{x}_k + \sum_{k=K_{NF}+1}^K \mathbf{H}_k \mathbf{x}_k + \mathbf{n} \end{aligned} \quad (11)$$

where $\mathbf{y} \in \mathbb{C}^{M \times 1}$ is the received signal vector, $\mathbf{H} \in \mathbb{C}^{M \times K \cdot N_U}$ is the channel matrix, $\mathbf{x} \in \mathbb{C}^{K \cdot N_U \times 1}$ is the transmitted symbol vector, and $\mathbf{n} \sim \mathcal{CN}(0, \sigma_n^2 \mathbf{I}_M)$ is the noise vector.

III. PROPOSED SIC MMSE RECEIVER WITH WEIGHTED-RATE (WR) ORDERING

In this section, we present the proposed SIC MMSE receiver with weighted-rate ordering that exploits the unique spatial characteristics of near-field XL-MIMO systems. We develop the mathematical framework for the weighted-rate metric computation, present the ordering algorithm, and analyze the interference cancellation process along with computational complexity.

A. Weighted-Rate Metric and MMSE Filter

In near-field XL-MIMO systems, the per-element distance computation in (4) produces nonlinear phase profiles across the array for each user. By contrast, the far-field model of (5)–(7) imposes a linear phase gradient, yielding a rank-1 or low-rank channel structure per user [2]. This nonlinear phase

diversity leads to richer singular value distributions and higher effective channel rank [1], [2], as explicitly demonstrated for ULA configurations in [20].

We first perform a singular value decomposition (SVD) of the channel matrix:

$$\mathbf{H} = \mathbf{U}\mathbf{\Sigma}\mathbf{V}^* \quad (12)$$

where $\mathbf{\Sigma} = \text{diag}(\sigma_1, \sigma_2, \dots, \sigma_{\min(M, K \cdot N_U)})$ with $\sigma_1 \geq \sigma_2 \geq \dots \geq \sigma_{\min(M, K \cdot N_U)}$.

The total spatial degrees of freedom are given by

$$\text{DoF} = \text{rank}(\mathbf{H}) \leq \min(M, K \cdot N_U) \quad (13)$$

Under the considered system setup, let $\mathcal{P}$ denote the uniform spatial distribution of user positions within the near-field region ($d < R$), modeled via (2)–(4), and at equivalent distances under the plane wave approximation of (5)–(7). Then, the expected effective rank of the channel matrix $\mathbf{H}$ satisfies:

$$\mathbb{E}_{\mathcal{P}}[\text{rank}(\mathbf{H})|_{\text{NF}}] \geq \mathbb{E}_{\mathcal{P}}[\text{rank}(\mathbf{H})|_{\text{FF}}] \quad (14)$$

where $\mathbb{E}_{\mathcal{P}}[\cdot]$ denotes expectation over $\mathcal{P}$, and the subscripts NF and FF indicate evaluation under the near-field and far-field channel models, respectively. This rank increase is due to the distance-dependent quadratic phase term $\pi n^2 d^2 \cos^2 \theta_{k,m} / (\lambda r_{k,m})$ in the near-field array response, which vanishes in the far field. Two users at the same angle but different distances have linearly independent near-field channel vectors, whereas their far-field channel vectors are collinear. This distance-domain separability raises the effective rank over the user distribution [1], [20].

The weighted-rate metric for the k -th user exploits the singular value structure from (12) and is defined by

$$M_k = \sum_{i=1}^{N_U} \sigma_{k,i}^2 \log_2 \left(1 + \frac{P_k \sigma_{k,i}^2}{\sigma_n^2} \right), \quad (15)$$

where $\sigma_{k,i}$ are the singular values of the user-specific channel matrix $\mathbf{H}_k$ obtained from:

$$\mathbf{H}_k = \mathbf{U}_k \mathbf{\Sigma}_k \mathbf{V}_k^* \quad (16)$$

with $\mathbf{\Sigma}_k = \text{diag}(\sigma_{k,1}, \sigma_{k,2}, \dots, \sigma_{k,N_U})$.

The optimal linear MMSE receive filter is well-known to be [4]:

$$\mathbf{W}_{\text{MMSE}} = \mathbf{H}^H (\mathbf{H}\mathbf{H}^H + \sigma_n^2 \mathbf{I}_M)^{-1} \quad (17)$$

B. Proposed Ordering and Interference Cancellation

The proposed ordering algorithm prioritizes users with the highest weighted-rate metric from (15). The metric values are computed for all users using their individual SVD from (16) and sorted to obtain the detection order as follows:

$$M_{k_1} \geq M_{k_2} \geq \dots \geq M_{k_K}, \quad (18)$$

yielding the detection order $\pi = [k_1, k_2, \dots, k_K]$.

The proposed SIC MMSE receiver with weighted-rate ordering involves the following steps. In the initialization, we set the received vector $\mathbf{y}$ as the vector for the first SIC stage and the set of $K \cdot N_U$ data streams given by

$$\mathbf{y}^{(1)} = \mathbf{y}, \quad \mathcal{R}^{(1)} = \{1, 2, \dots, K \cdot N_U\} \quad (19)$$

For the $K \cdot N_U$ data streams, we perform user selection based on the weighted-rate ordering in (18):

$$k_i^* = \pi(i) \quad (20)$$

For each user k_i^* , individual streams are ordered by descending singular values $\sigma_{k_i^*,1} \geq \sigma_{k_i^*,2} \geq \dots \geq \sigma_{k_i^*,N_U}$ from the user-specific SVD in (16).

The MMSE receive filter for this stream is applied to the residual received signal $\mathbf{y}^{(i)}$ to obtain the soft estimate:

$$\bar{x}_{k_i^*} = \mathbf{w}_{k_i^*}^H \mathbf{y}^{(i)} \quad (21)$$

where $\mathbf{w}_{k_i^*}$ is the corresponding row of $\mathbf{W}_{\text{MMSE}}$ in (17). The soft estimate is then mapped to the nearest constellation point by a symbol-by-symbol minimum-distance decision:

$$\hat{x}_{k_i^*} = \mathcal{Q}(\bar{x}_{k_i^*}) \quad (22)$$

where $\mathcal{Q}(\cdot)$ denotes the hard-decision quantizer that maps the MMSE soft estimate to the nearest constellation symbol. The contribution of the detected symbol is then subtracted from the residual:

$$\mathbf{y}^{(i+1)} = \mathbf{y}^{(i)} - \mathbf{h}_{k_i^*} \hat{x}_{k_i^*} \quad (23)$$

In the last step of each SIC stage, we remove the detected stream from the remaining set of data streams:

$$\mathcal{R}^{(i+1)} = \mathcal{R}^{(i)} \setminus \{i\} \quad (24)$$

Unlike channel-norm ordering, which is rate-blind, and capacity-based ordering, which weights all streams equally, the proposed metric weights each stream's rate $\log_2(1 + P_k \sigma_{k,i}^2 / \sigma_n^2)$ by its squared singular value $\sigma_{k,i}^2$, prioritizing streams that are both reliable to cancel first and information-rich, which limits error propagation in SIC [1], [2]. The computational cost of the proposed and existing linear and SIC receivers is shown in TABLE I.

TABLE I: Computational Cost Comparison

Detection Scheme	Complexity
Linear MMSE	$\mathcal{O}(\min(M^3, (KN_U)^3))$
Channel Norm SIC [4]	$\mathcal{O}(K^2 \cdot M \cdot N_U)$
SINR-based SIC [6]	$\mathcal{O}(K^2 \cdot M \cdot N_U)$
Proposed WR-SIC	$\mathcal{O}(M^3 + K \cdot M \cdot N_U^2)$

C. Sum-Rate Analysis

The sum-rate performance of the proposed weighted-rate SIC receiver can be obtained by assuming Gaussian signaling and considering the effective SINR after each stage. Following the standard turbo-IDD analysis methodology [11], [12], the Gaussian signaling assumption provides a theoretical upper bound on the achievable rate per stream, while the simulations in Section V employ QPSK with LDPC coding operating below this bound.

For the i -th detected stream, the achievable rate is given by

$$R_i = \log_2(1 + \text{SINR}_i^{\text{eff}}) \quad (25)$$

The effective SINR for the i -th stream after SIC is:

$$\text{SINR}_i^{\text{eff}} = \frac{P_{k_i^*} |\mathbf{w}_{k_i^*}^H \mathbf{h}_{k_i^*}|^2}{\sum_{j \in \mathcal{R}^{(i+1)}} P_j |\mathbf{w}_{k_i^*}^H \mathbf{h}_j|^2 + \sigma_n^2 \|\mathbf{w}_{k_i^*}\|^2} \quad (26)$$

where $\mathbf{w}_{k_i^*}$ is the MMSE filter for the k_i^* -th user from (17), and $\mathcal{R}^{(i+1)}$ is the remaining stream set from (24). The total achievable sum-rate for the system is:

$$R_{\text{sum}} = \sum_{i=1}^{K \cdot N_U} R_i = \sum_{i=1}^{K \cdot N_U} \log_2 (1 + \text{SINR}_i^{\text{eff}}) \quad (27)$$

For near-field scenarios, the enhanced spatial degrees of freedom from (14) lead to improved sum-rate performance:

$$\mathbb{E}_{\mathcal{P}}[R_{\text{sum}}^{NF}] \geq \mathbb{E}_{\mathcal{P}}[R_{\text{sum}}^{FF}] \quad (28)$$

This is achieved by the weighted-rate ordering from (15), which prioritizes users with better channel conditions, reducing error propagation and maximizing system throughput.

IV. PROPOSED ITERATIVE DETECTION AND DECODING

The proposed IDD integrates WR-SIC ordering with LDPC coding following [11], [12], but differs from [11] (which addresses 2x2 block-fading MIMO with no ordering) by operating on near-field XL-MIMO LoS channels with spherical wavefront geometry (2)–(4), incorporating M_k from (15) in each iteration, and targeting hybrid near/far-field deployments. A rate 1/2 LDPC code with log-domain belief propagation is employed, exchanging Log-Likelihood Ratios (LLRs) between detector and decoder until convergence.

A. Scalar Equivalent Model and LLR Computation

Each data stream is modeled using a scalar Gaussian approximation:

$$\tilde{x}_{\pi(i)} = \mu_{\pi(i)} x_{\pi(i)} + \xi_{\pi(i)} \quad (29)$$

$$\tilde{x}_{\pi(i)} = \text{LDPC-Decoder}(\text{LLR}(\tilde{x}_{\pi(i)})) \quad (30)$$

where $\tilde{x}_{\pi(i)}$ represents the output of the receive filter before LDPC decoding, $\mu_{\pi(i)} = \mathbf{w}_{\pi(i)}^H \mathbf{h}_{\pi(i)}$ is the effective gain, and $\xi_{\pi(i)} \sim \mathcal{CN}(0, \sigma_{\pi(i)}^2)$ represents residual interference plus noise. The refined estimates $\tilde{x}_{\pi(i)}$ obtained from the LDPC decoder output are used for interference cancellation in subsequent IDD iterations.

The log-likelihood ratio (LLR) for bit b_l of symbol x_k is:

$$L_D(b_l) = \log \frac{\sum_{x \in \mathcal{X}_l^{+1}} \exp\left(-\frac{|\tilde{x}_k - \mu_k x|^2}{\sigma_k^2}\right) P(x)}{\sum_{x \in \mathcal{X}_l^{-1}} \exp\left(-\frac{|\tilde{x}_k - \mu_k x|^2}{\sigma_k^2}\right) P(x)} - L_C(b_l) \quad (31)$$

where $\mathcal{X}_l^{+1}$ and $\mathcal{X}_l^{-1}$ are constellation symbol sets with bit $b_l = +1$ and $b_l = -1$, and $L_C(b_l)$ is the extrinsic LLR from the LDPC decoder.

The a priori probability of symbol x is:

$$P(x) = \prod_{m=1}^{M_c} \frac{1}{1 + \exp(-x_m^b L_C(b_m))} \quad (32)$$

where M_c is the number of bits per symbol.

The residual variance after SIC processing is:

$$\sigma_k^2 = \sigma_n^2 \|\mathbf{w}_k\|^2 + \sum_{j \neq k} \text{Var}(x_j) |\mathbf{w}_k^H \mathbf{h}_j|^2 \quad (33)$$

B. Message Passing Algorithm

For each iteration t , the detector computes extrinsic LLRs:

$$L_D^{(t)}(b_l) = \log \frac{\sum_{x \in \mathcal{X}_l^{+1}} \exp\left(-\frac{|\tilde{x}_k^{(t)} - \mu_k^{(t)} x|^2}{\sigma_{k,t}^2}\right) P^{(t-1)}(x)}{\sum_{x \in \mathcal{X}_l^{-1}} \exp\left(-\frac{|\tilde{x}_k^{(t)} - \mu_k^{(t)} x|^2}{\sigma_{k,t}^2}\right) P^{(t-1)}(x)} \quad (34)$$

The LDPC decoder updates a priori probabilities as:

$$P^{(t)}(x) = \prod_{m=1}^{M_c} \frac{\exp(x_m^b L_C^{(t)}(b_m)/2)}{2 \cosh(L_C^{(t)}(b_m)/2)} \quad (35)$$

The process continues until the convergence of the LDPC decoder, the maximum number of iterations, or the LLR threshold is reached.

V. NUMERICAL RESULTS

The evaluation of the proposed and existing interference cancellation schemes is conducted through Monte Carlo simulations. The simulation parameters are carefully selected to represent realistic near-field XL-MIMO scenarios while enabling direct comparison between different techniques. We assume that the interference cancellation schemes have perfect channel state information. Table II gives the simulation parameters.

TABLE II: Simulation Parameters

Parameter	Value
Number of BS receive antennas (M)	64
Number of users (K)	16
Transmit antennas per user (N_U)	2
Total spatial streams ($K \times N_U$)	32
Antenna spacing (d)	$\lambda/2$
Modulation scheme	QPSK
Channel coding	LDPC (rate 1/2)
Maximum IDD iterations	3
LDPC decoder iterations	20

For the hybrid scenario, half the users are randomly placed in the near-field region ($d < R$) and half in the far-field ($d \geq R$), with distances drawn uniformly from $[1.2D, R]$ and $[R, 3R]$, respectively [3]. Each simulation point averages at least 10^4 independent channel realizations. For low BER points, additional codewords are transmitted over the same 10^4 realizations to reduce variance.

The results show superior performance of the proposed WR-SIC over conventional detection schemes across all scenarios. The left plot of Fig. 1 shows the average BER performance comparison between near-field and far-field users for all detection techniques. Near-field users (solid) outperform far-field users (dashed), with the gap increasing at higher SNR due to spherical wavefront spatial resolvability. The proposed WR-SIC achieves approximately 2–3 dB SNR gain compared to channel norm [4] and SINR-based SIC [6] at BER of 10^{-3} , with linear MMSE exhibiting the poorest performance across both user categories.

The right plot of Fig. 1 presents the sum-rate performance comparison between near-field and far-field users using different detection schemes. The results corroborate the BER findings, showing that near-field users consistently achieve higher

sum-rates than their far-field counterparts. The proposed WR-SIC demonstrates the highest sum-rate performance for both types of users, exploiting the enhanced spatial discrimination capabilities of near-field propagation.

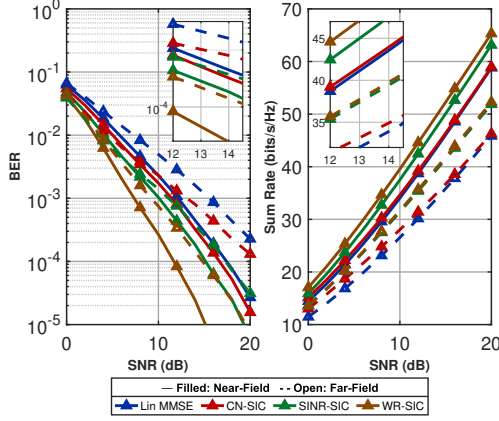


Fig. 1: BER and sum rate performance comparison of near and far field users under perfect CSI

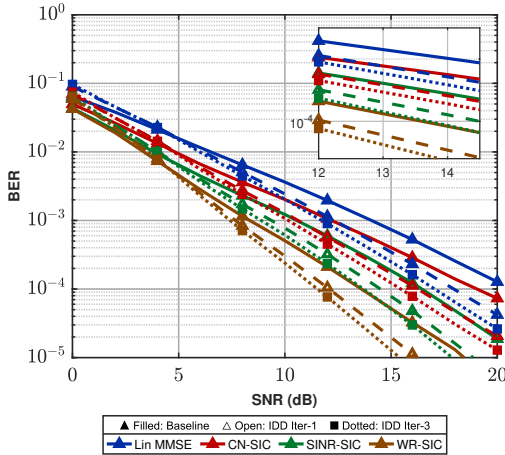


Fig. 2: Average BER over all users (half near-field, half far-field) with and without IDD using 1 and 3 iterations

Fig. 2 reports the average BER over all $K = 16$ users, providing a system-level view of IDD gains across the full hybrid user population. All detection schemes benefit from iterative processing, with the proposed WR-SIC achieving approximately 1–2 dB additional SNR gain at BER of 10^{-3} with just 1 iteration. Extending to 3 IDD iterations provides further enhancements across all methods. The diminishing returns with extra iterations suggest that most gains are captured in the first iterations, making 3 iterations practical.

VI. CONCLUSION

This work presented an interference cancellation strategy combining a novel weighted-rate SIC ordering technique and an IDD scheme for LoS channels. Unlike conventional channel norm or SINR-based ordering, the proposed method exploits enhanced spatial degrees of freedom through weighted-rate analysis that leverages the geometric properties of spherical wavefront propagation. The results show superior performance

of near-field users to far-field users, validating the effectiveness of the proposed approach in XL-MIMO systems.

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